

An Adjustable RF Tuning Element for Microwave, Millimeter wave, and Submillimeter Wave Integrated Circuits

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ABSTRACT

Planar RF circuits are used in a wide range of applications from 1 GHz to 300 GHz including radar, communications, commercial RF test instruments, and remote sensing radiometers. These circuits, however, provide only fixed tuning elements. This lack of adjustability puts severe demands on circuit design procedures and materials parameters. We have developed a novel tuning element which can be incorporated into the design of a planar circuit in order to allow active, post-fabrication tuning by varying the electrical length of a coplanar strip transmission line. It consists of a series of thin plates which can slide in unison along the transmission line, and the size and spacing of the plates are designed to provide a large reflection of RF power over a useful frequency bandwidth. Tests of this structure at 1 GHz to 3 GHz showed that it produced a reflection coefficient greater than 0.90 over a 20% bandwidth. A 2 GHz circuit incorporating this tuning element was also tested to demonstrate practical tuning ranges. This structure can be fabricated for frequencies as high as 1000 GHz using existing micromachining techniques. Many commercial applications can profit from this micromechanical RF tuning element, as it will aid in extending microwave integrated circuit technology into the high millimeter wave and submillimeter wave bands by easing constraints on circuit technology.

INTRODUCTION

The vast field of RF electronics has long capitalized on the advantages of simple and cost effective planar circuit technology. From 1 GHz to 40 GHz, microstrip, coplanar strip, and slotted transmission lines are commonly used for signal distribution in hybrid and monolithic circuits used in radar, communications, and remote sensing applications. Typically, the transmission lines in these circuits are designed with a particular characteristic impedance, and some form of impedance transformation or tuning circuit is used to insure that the circuit components transfer the RF signal along these lines in the most efficient way. Ideally, knowledge of the operating impedance of each component allows for a fixed circuit design which incorporates this impedance transformation. In practice, however, it is quite difficult to accurately characterize these components, particularly at high frequencies. As a result, some post fabrication tuning in the form of circuit modification is required. This type of tuning is not easy and it is required more often as the design frequency is increased. It is therefore desirable to incorporate impedance matching elements in these circuits which can be readily finely tuned after fabrication in order to relax the constraint of component characterization, and thus allow the circuit design to be extended to higher frequencies.

In waveguide circuits, this tuning is accomplished with a mechanically adjustable backshort which is inserted into the waveguide. This provides a tuning stub with an adjustable electrical length. Waveguide, however, is often difficult to interface with planar components, and the dimensional requirements can make the waveguide circuit quite difficult and costly to fabricate. Analogously, an approach for a movable, noncontacting "planar backshort" which can be used to vary the electrical length of a coplanar strip transmission line tuning stub has been developed. This sliding circuit element allows for the design of an RF circuit which can be actively tuned after fabrication to achieve optimal performance.

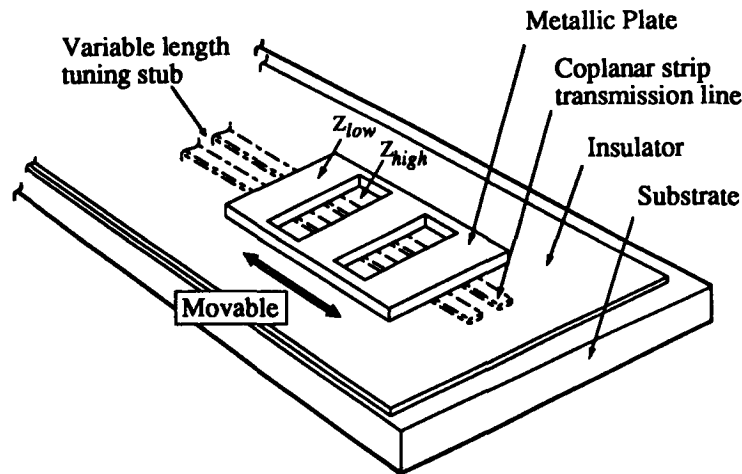


Fig. 1. Design of sliding backshort on coplanar strip transmission line. The holes in the metal plate create a series of successive high and low impedance sections that produce a large reflection of RF power.

THE PLANAR BACKSHORT

Placing a solid metallic plate across a coplanar strip transmission line, with a thin dielectric insulating layer in between, results in a reflection of RF power. Unfortunately, this reflection is not sufficiently large over a useful bandwidth for the plate to be used as a backshort in the design of a practical tuning circuit. This “sandwich” configuration does, however, result in a section of lower impedance transmission line. Quarter-wavelength sections of this line can be cascaded alternately with uncovered sections of “high” impedance line to create a series of impedance transformations which ultimately result in a very low impedance. When used as a backshort on a transmission line, this cascade produces an even larger reflection of RF power than the solid plate. We constructed and optimized such a planar backshort and its geometry is illustrated in Fig. 1. It consists of a thin metal plate with rectangular holes of the proper dimension and spacing. A thin insulator keeps the plate from contacting the transmission line which reduces wear and allows the backshort to slide freely when pushed.

MODEL MEASUREMENTS

We built a large scale model of the sliding backshort and measured the magnitude of the reflection coefficient with an HP 8510B network analyzer over a frequency range of 1 GHz to 5 GHz. This measurement required a transition between the coaxial input of the network analyzer and the coplanar line. Unfortunately, no standard transition with low VSWR is readily available. For this reason, we devised several distinct measurement techniques in order to obtain reliable results.

A natural choice of calibration technique for an HP8510B is the “Thru-Reflect-Line” (TRL) method which allows for measurements in nonstandard transmission media such as coplanar line. The transition between coaxial and coplanar line can, in principle, be accounted for in the calibration procedure. However, reproducible calibration standards in coplanar line are required. Reflections at connections between segments of coplanar line and uncertainties in the reflection standard lead to nonrepeatable results and unacceptably high errors. As a result, this method was not pursued further.

Our second technique was a simple 1-port measurement of a circuit which employs a direct connection from coaxial to coplanar line. This measurement of $|s_{11}|$ was made over a 1.5 GHz to 2.5 GHz frequency range and Fig. 2(a) shows the test arrangement. The abrupt transition from coaxial line to coplanar line was formed at the edge of the stycast substrate with a flange mount SMA connector. The measurement was taken with the reference plane of the backshort adjusted to coincide with the SMA connector.

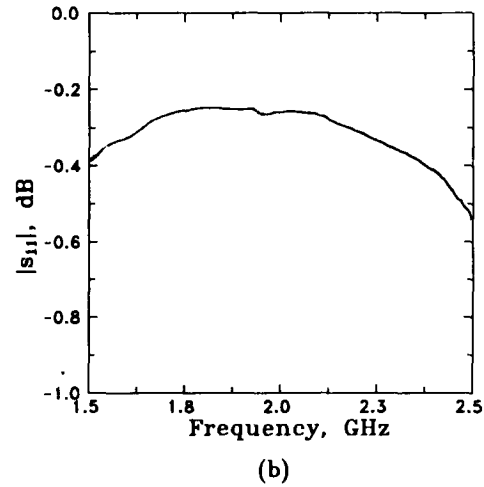
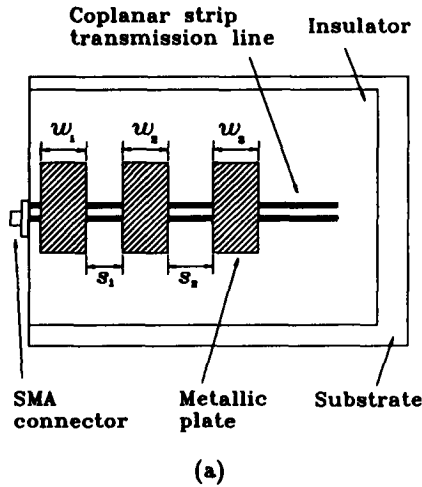


Fig. 2. Test circuit used for optimization (a) and plot of reflection coefficient (b) for optimized sliding backshort at the SMA connector on 204- Ω transmission line.

While this transition resulted in large unwanted reflections which increased the uncertainty of the measurement, the technique was useful because it allowed us to monitor the reflection coefficient for the backshort in real time as we optimized the dimensions of the backshort. The optimization was performed by systematically varying the length of the low and high impedance sections, the number of sections, the dimensions of the transmission line, and the thickness of the insulator in order to achieve the largest reflection of RF power. Good performance was obtained for a coplanar transmission line with 2.1 mm wide copper strips, separated by a 5.2 mm gap and mounted on a 6 mm thick styrocast substrate with a dielectric constant of 4. The characteristic impedance of this transmission line and its effective dielectric constant were determined to be 204 Ω and 2.3 respectively [1]. A 0.025 mm thick sheet of mylar was used to insulate the transmission line from the sliding shorting plate. This noncontacting, 76 mm wide, 6 mm thick aluminum shorting plate had two rectangular holes in it with dimensions and spacing of $w_1 = 24.3$ mm, $s_1 = 19.4$ mm, $w_2 = 24.0$ mm, $s_2 = 23.0$ mm, and $w_3 = 24.4$ mm. This resulted in uncovered high impedance sections, s_1 and s_2 , and covered low impedance sections, w_1 , w_2 and w_3 , which were each approximately $\lambda_g/4$ long on the coplanar line.

A plot of $|s_{11}|$ versus frequency is shown in Fig. 2(b). This optimized planar backshort produced $|s_{11}|$ better than -0.3 dB over a 20% bandwidth. That is a reflection of more than 90% of the power in the incident wave. The center frequency was determined to be 2 GHz, which agrees exactly with the design frequency.

We noted that the front edge of the sliding metallic plate was very close to the discontinuity at the SMA connector and it may have interacted with fringing fields around the flange. In addition, the reflection coefficient decreased when the backshort was moved $\lambda_g/2$ from the SMA connector in order to reduce interactions with fringing fields. The error caused by this unwanted interference, along with the lack of a transmission measurement, motivated the design of a third measurement technique.

Fig. 3 shows the system used for the third measurement technique. Here, the 204- Ω coplanar line was connected to the 50- Ω network analyzer inputs by means of two baluns of identical length. These baluns were made by gradually trimming the shield and teflon insulation from a semirigid, 3.5 mm wide coaxial line over approximately one wavelength at 2 GHz. This created a smooth transition to the coplanar line which minimized the power reflected at the connection. The return loss for these baluns is approximately -10 dB and the remaining undesired reflections from these transitions were gated out of the measurement using the low-pass time domain mode of the network analyzer. The frequency for this measurement was swept from 50 MHz to 20 GHz so that an accurate transformation between frequency and time could be made.

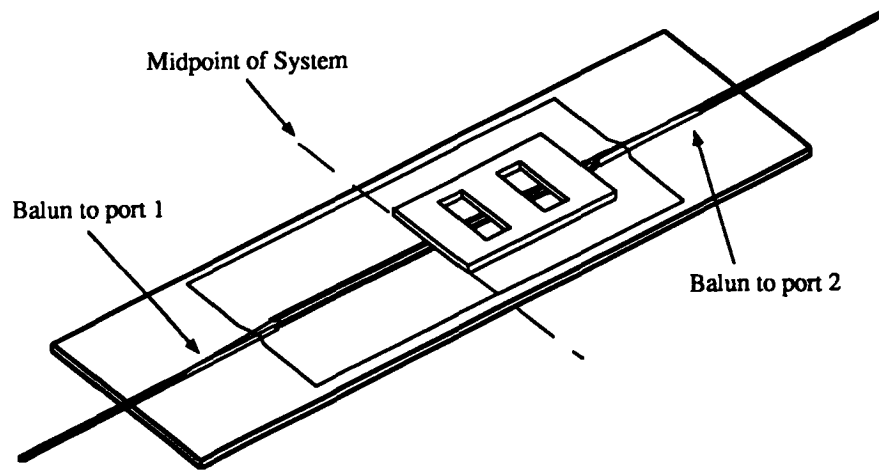


Fig. 3. Test system used to measure two-port scattering parameters for sliding backshort. The coaxial line baluns are tapered to create a gradual transition between coaxial and coplanar line which reduces measurement uncertainty due to unwanted reflections at the transition.

The full two-port scattering parameters for the system were measured under three different conditions. First, a reference measurement was made which would correspond to an ideal short. The baluns are identical in length and hence, the test model is symmetric about the midpoint of the coplanar line. Thus, the magnitude of the transmission measurement of this circuit with no short in place is equal to the reflection measurement with an ideal short at the midpoint. Reflection measurements for the sliding backshort were then made with the shorting plate arranged to reflect an incident wave from port 1 at the midpoint of the system and then, port 2. The values for $|s_{11}|$ from the first reflection measurement and $|s_{22}|$ from the second were normalized by dividing each by the values for $|s_{21}|$ and $|s_{12}|$ from the reference measurement, respectively. The two results were averaged to cancel the effect of any asymmetry in the system. Transmission measurements for the backshort were similarly normalized and averaged. The requirement of processing the measurement data, along with the large range of frequency, prohibited us from monitoring the 2 GHz reflection coefficient in real time for this measurement.

The results for the averaged, normalized $|s_{11}|$ and $|s_{21}|$ are shown in Fig. 4. The plot shows that $|s_{11}|$ was

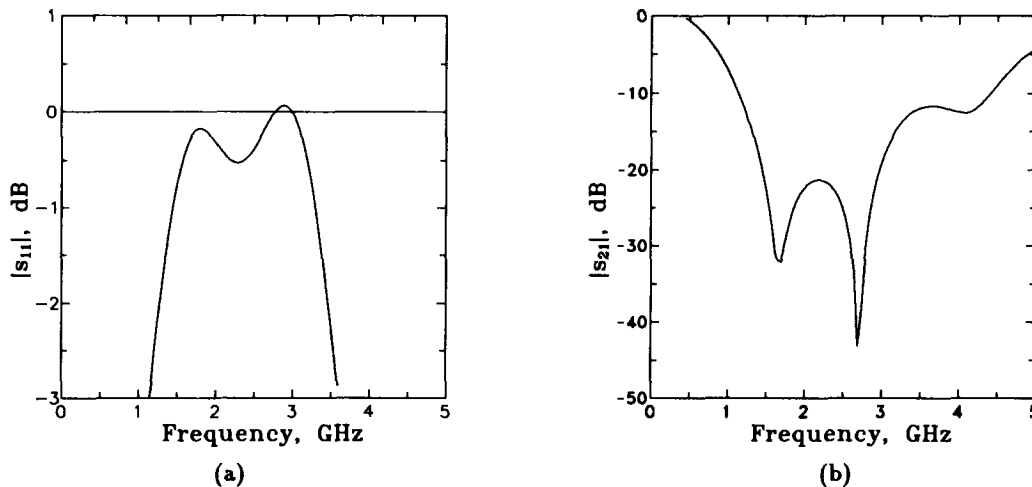


Fig. 4. Plot of measured reflection coefficient (a) and transmission coefficient (b) for optimized sliding backshort. This measurement was made using baluns for transitions.

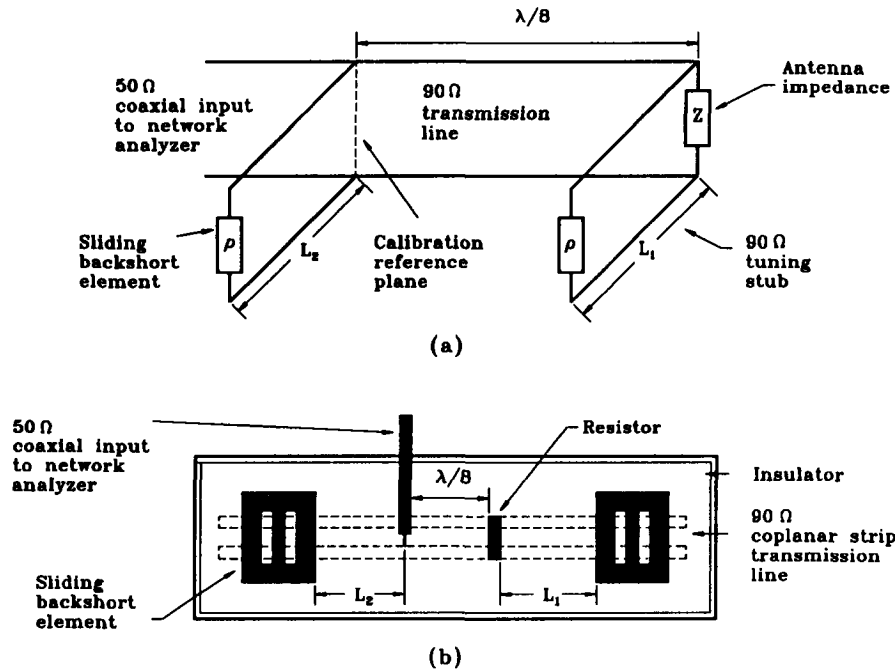


Fig. 5. Schematic diagram for equivalent double-shunt stub tuner circuit (a) and circuit arrangement used for measurements (b).

better than -0.5 dB over approximately an 80% bandwidth and the center frequency was slightly higher than 2 GHz. Over the peak located between 1.5 GHz and 2.5 GHz, $|s_{11}|$ is better than -0.3 dB over a 16% bandwidth and the center frequency is slightly lower than 2 GHz. This agrees well with the previous results shown in Fig. 2(b). Together, the plots of $|s_{11}|$ and $|s_{21}|$ appear to indicate that approximately 10% of the power for the incident wave is left unaccounted for, but the ± 0.2 dB uncertainty of our measurement is too large to verify this.

DOUBLE STUB TUNER

In order to demonstrate the tuning range accessible in a planar circuit, we built a double-shunt stub tuner which incorporates two sliding backshorts. This circuit serves as a low frequency model of a superconductor-insulator-superconductor (SIS) mixer used in millimeter wave radioastronomy applications [2,3]. Fig. 5(a) shows the equivalent circuit and Fig. 5(b) shows the circuit arrangement as measured. The characteristic impedance of the coplanar line is 90Ω and the stub spacing is $\lambda_g/8$. A $90\text{-}\Omega$ resistor was used to simulate a planar antenna impedance and a 3.5 mm wide semirigid coaxial probe was used to measure the range of impedance to which we could transform the resistor. The calibration reference plane for this measurement was set at end of the shield for the coaxial probe. Fig. 6 is a Smith chart, normalized to the characteristic impedance of the line, which shows the accessible impedance region at 2 GHz. The overlap for the tuning region and the impedance region necessary for matching SIS devices implies that a circuit of this type would be useful for this purpose. Variations in the shape of this tuning region can be achieved by changing the spacing between the tuning stubs.

The solid and dashed boundary lines in Fig. 6 show a comparison between the measured impedance range of the double stub tuner and a computer simulation of the circuit using *Puff* [4], respectively. The simulation agrees well with the measured data. The reflection coefficient of the shorting element used in the computer model was adjusted to fit the simulation to the measured Smith chart data. The resulting fitted $|s_{11}|$ for the backshort was -1.0 dB, which is similar to the -0.7 dB result that was measured for the

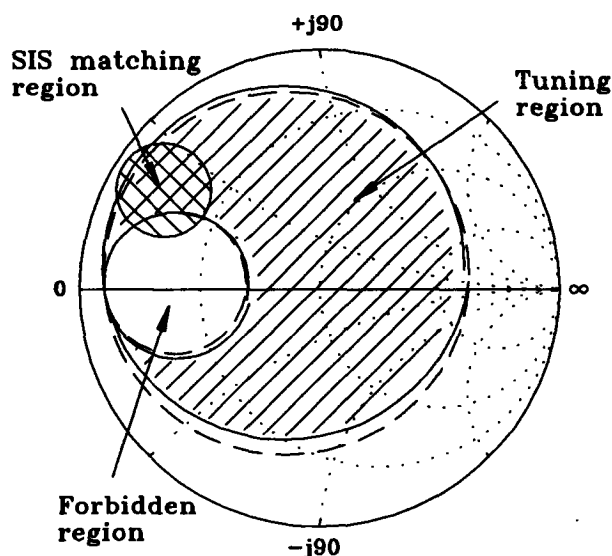


Fig. 6. Smith chart showing the measured (solid boundary) and fitted (dashed boundary) tuning region for double shunt stub tuner. The available tuning region covers the impedance region needed for matching to an SIS device.

same backshort, by itself, on a $90\text{-}\Omega$ transmission line at 2 GHz. A phase shift of 4 degrees was also fitted at the coaxial probe transition in order to account for the calibration reference plane uncertainty.

CONCLUSION

We have demonstrated an approach for an adjustable planar backshort on coplanar strip transmission line. Test results from a low-frequency model indicate that the backshort can be used to create tuning stubs whose electrical length can be varied after fabrication. This noncontacting backshort with cascaded high and low impedance sections should also work on slotline, coplanar waveguide and possibly microstrip line. By using advanced micro-machining techniques [5,6], it should be possible to create adjustable impedance matching circuits at terahertz frequencies which would relax the design constraints for a wide range of planar integrated circuits.

ACKNOWLEDGEMENTS

We wish to thank Y-C. Tai for his help in keeping the design viable for terahertz scaling through micro-machining techniques. We also wish to thank O. Borić, M. A. Frerking, E. Kollberg, K. Potter, P. Siegel, and T. Tolmunen for valuable discussions. This work was supported in part by the Jet Propulsion Laboratory, California Institute of Technology, under contract with the National Aeronautics and Space Administration and the Innovative Science and Technology Office of the Strategic Defense Initiative Organization.

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